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Final Report

Signal Structure Methods for Robust Interference Reduction in Airborne Antenna Arrays

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1 Introduction and Summary

Tactical airborne communications intelligence (COMINT) collection platforms are increasingly faced with multiple spectrally-overlapped signals. Direction finding (DF) and interference reduction for signals of interest in such environments pose significant technical challenges and are a high-priority need for the armed forces. Under this contract we have developed a variety of innovative signal processing techniques for DF and interference reduction.

Airborne collection platforms usually operate over friendly territory at a stand-off range of 50–100 miles from the forward line of troops (FLOT). The platform flies between altitudes of 20,000 to 70,000 feet in order to receive ("listen" to) SOIs across the line of battle. At 30,000 feet, an airborne platform can receive signals from over 200 miles. Co-channel interference arises from several sources. First, tactical wireless networks re-use frequency every 10 to 20 miles by exploiting high ground-to-ground propagation loss. However, the ground-to-air propagation loss to the airborne collection platform can be much lower causing significant co-channel interference for the collection platform. Second, friendly radios which are much nearer to airborne platform, sometimes operate in the same frequency bands as the SOIs, creating severe co-channel interference. Next, SOIs can sometimes intentionally hide near powerful commercial signals (TV and broadcast FM) to mask their presence. Finally, jammers may be deployed to prevent collection of high value intelligence signals. All these conditions can often result in up to four or five signals with the SOIs buried 10–30 dB below the other co-channel signals.

A promising approach to this problems is to use space-time (ST) signal processing for equalization and co-channel interference reduction. We have developed new techniques in the area of blind channel estimation and equalization. The techniques make use of the structure of the received signals. Subspace approaches can be used to exploit the spatial structure of the signals. ST techniques can extend subspace methods to the optimum beamformers. Additionally, constant modulus schemes as well as higher- and second-order statistics can exploit the temporal structure in the signals and have been recently proposed for blind channel estimation and equalization. The performance of blind methods will, of course, be sensitive to the validity of structural properties assumed.

2 Signal Model

Let the continuous-time output from the receive antenna array $\mathbf{x}(t)$ be sampled at the symbol rate at instants $t = t_0 + kT$.

Then the output may be written as

$$\mathbf{x}(k) = \mathbf{H}\mathbf{s}(k) + \mathbf{n}(k) \tag{1}$$

where $\mathbf{H}$ is the symbol response channel (a $m \times N$ matrix) that captures the effects of the array response, symbol waveform and path fading. m is the number of antennas, N is the channel length in symbol periods and $\mathbf{n}(k)$ the sampled vector of additive noise. Note that $\mathbf{n}(k)$ may be colored in space and time, as will be shown later. $\mathbf{H}$ is assumed to be time invariant. i.e., α^R is constant. $\mathbf{s}(k)$ is a vector of N consecutive elements of the data sequence and is defined as

$$\mathbf{s}(k) = \begin{bmatrix} s(k) \\ \vdots \\ s(k-N+1) \end{bmatrix} \tag{2}$$

Note that GSM uses GMSK modulation where the transmitted signal is not a linear map of the underlying data sequence. In such a case, the signal model in Eq. (1) cannot be strictly used. In practice, a reasonable linear approximation of GSMK modulation is possible and the model described above remains applicable.

2.1 Block signal model

It is often convenient to handle signals in blocks. Therefore we may collect M consecutive snapshots of $\mathbf{x}(\cdot)$ corresponding to time instants $k, \ldots, k+M-1$, we get

$$\mathbf{X}(k) = \mathbf{H}\mathbf{S}(k) + \mathbf{N}(k) \tag{3}$$

where $\mathbf{X}(k)$, $\mathbf{S}(k)$ and $\mathbf{N}(k)$ are defined as

$$\mathbf{X}(k) = [\mathbf{x}(k) \cdots \mathbf{x}(k+M-1)] \qquad (m \times M)$$

$$\mathbf{S}(k) = [\mathbf{s}(k) \cdots \mathbf{s}(k+M-1)] \qquad (N \times M)$$

$$\mathbf{N}(k) = [\mathbf{n}(k) \cdots \mathbf{n}(k+M-1)] \qquad (m \times M)$$

Note that $\mathbf{S}(k)$ by definition is constant along the diagonals and is therefore Toeplitz.

Given the signal model at Eq. (3), an important question is whether the unknown channel $\mathbf{H}$ and the data $\mathbf{s}$ can be determined from the observations $\mathbf{X}$. This leads us to examine the underlying constraints on $\mathbf{H}$ and $\mathbf{S}(\cdot)$ which we call *structure.*

2.2 Structure

Spatial structure It can be shown that the columns of $\mathbf{H}$ lie in a space spanned by $\mathbf{a}(\theta)$ which lie on the *array manifold* $\mathcal{A}$. This array manifold is a set of array response vectors indexed by θ

$$\mathcal{A} = \{\mathbf{a}(\theta) | \theta \in \Theta\} \tag{4}$$

where Θ is the set of all possible values of θ. Knowledge of $\mathcal{A}$ helps determine $\mathbf{a}(\theta_l)$. $\mathcal{A}$ includes the effect of array geometry, element patterns, inter-element coupling, scattering from support structures and objects near the base station. $\mathcal{A}$, when measured at the receiver baseband after digitization, includes the effects by cable and receiver gain/phase response, I-Q imbalance, A/D converter errors. $\mathcal{A}$ is frequency dependent and needs to be calibrated at multiple points within the operating band.

Temporal structure The temporal structure relates to the properties of the signal $u(t)$ and includes modulation format, pulse-shaping function and symbol constellation. Some typical temporal structures are

- Constant modulus (CM)

In many wireless applications, the transmitted waveform has a constant envelope (e.g., in FM modulation). A typical example of a constant envelope waveform is the GMSK modulation used in the GSM cellular system which has the following general form

$$u(t) = e^{j(\omega t+\phi(t))}$$

where $\phi(t)$ is a Gaussian-filtered phase output of a minimum shift keyed (MSK) signal.

- Finite alphabet (FA)

Another important temporal structure in mobile communication signals is the *finite alphabet.* This structure underlies all digitally modulated schemes. The modulated signal is a linear or nonlinear map of an underlying finite alphabet. For example, the IS-54 signal is a $\pi/4$ shifted DQPSK signal given by

$$u(t) = \sum_p A_p g(t-pT) + j\sum_p B_p g(t-pT) \qquad (5)$$

$$A_p = cos(\phi_p) \quad B_p = sin(\phi_p) \quad \phi_p = \phi_{p-1} + \Delta\phi_p$$

where $g(\cdot)$ is the pulse shaping function (which is a square root raised cosine function in the case of IS-54), and $\Delta\phi_p$ is chosen from a set of finite phase shifts $\{\frac{5\pi}{4}, \frac{3\pi}{4}, \frac{\pi}{4}, \frac{7\pi}{4}\}$ depending on the data $s(\cdot)$. This finite set of phase shifts represents the FA structure.

- Distance from Gaussianity

The distribution of digitally modulated signals is not Gaussian[1], and this property can be exploited to estimate the channel from higher-order moments such as cumulants. See e.g.

[1]The distribution may however approach Gaussian when constellation shaping is used for spectral efficiency [1]

[2], [3]. Clearly CM signals are non-Gaussian. These higher order statistics (HOS) based methods are usually slower converging than those based on second order statistics.

- Cyclostationarity

It can be shown that the continuous-time stochastic process $x(t)$ (assuming the fade amplitude α^R is constant) is cyclostationary. Moreover, the discrete sequence $\{x_i\}$ obtained by sampling $x(t)$ at the symbol rate $\frac{1}{T}$ is wide-sense stationary, whereas the sequence obtained by temporal oversampling (i.e., at a rate higher than $1/T$) or spatial oversampling (multiple antenna elements) is cyclostationary. The cyclostationary signal consists of a number of *phases* each of which is stationary. A phase corresponds to a shift in the sampling point in temporal oversampling and different antenna element in spatial oversampling. The duality between temporal and spatial oversampling is illustrated in Figure 1.

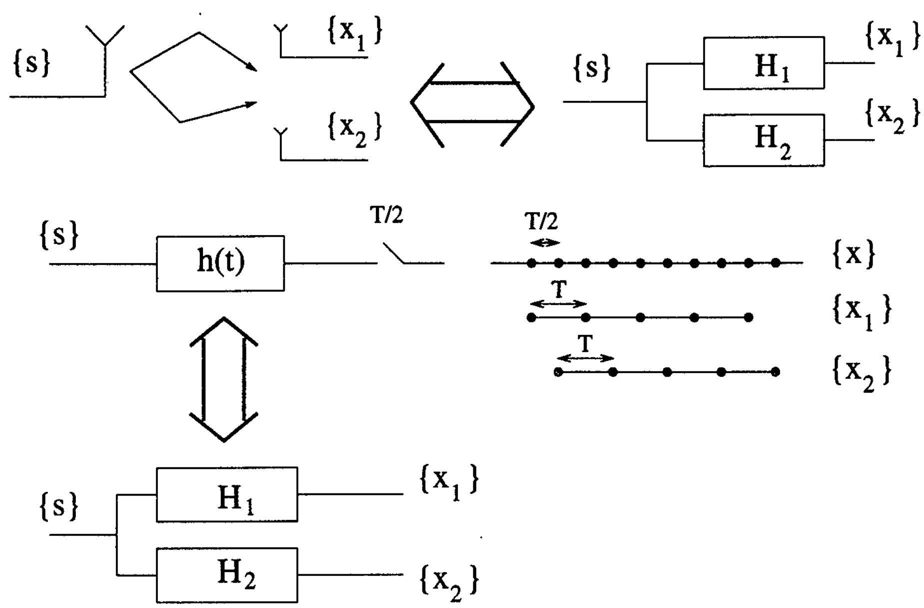

Figure 1: Antennas and/or oversampling result in polyphase SIMO channels

The cyclostationary property of sampled communication signals carries important information about the channel phase, which can be exploited in several ways to identify the channel. The cyclostationarity property can also be interpreted as a *finite duration* property. Put simply, this says that the oversampling increases the number of samples in the signal $\mathbf{x}(t)$ and phases in the the channel $\mathbf{H}$ but does not change the value of the data for the

duration of the symbol period. This allows $\mathbf{H}$ to become tall (more rows than columns) and full column rank. Also the stationarity of the channel makes $\mathbf{H}$ Toeplitz (or rather block Toeplitz). Tallness and Toeplitz properties are key to the blind estimation of $\mathbf{H}$.

- The Temporal Manifold

Just as the array manifold captures spatial wavefront information, the *temporal manifold* captures the temporal pulse-shaping function information (see [4], [5]). We define the temporal manifold $\mathbf{k}(\tau)$ as the sampled response of a receiver to an incoming pulse with delay τ. Unlike the array manifold, the temporal manifold can be estimated with good accuracy since it depends only on our knowledge of the pulse-shaping function. The following table shows the duality between the array and the temporal manifold.

Manifold	Indexed by	Characterizes
Array	angle θ	antenna array response
Time	delay τ	transmitted pulse shape

Table 1: The duality between the array and the time manifold

The different structures and properties inherent in the nature of the transmitted signals and the employed receivers in space-time processing are shown in Figure 2.

3 Blind Algorithms

The term "blind" methods (other names are "self-recovering" or "unsupervised"), do not need training signals and rather exploit the temporal structure such as non-Gaussianity, constant modulus (CM), finite alphabet (FA), cyclostationarity, or the spatial structure (such as the array manifold). The performance of blind methods will, of course, be sensitive to the validity of structural properties assumed.

Spatial structure or DOA based methods A *subspace approach* can be used to estimate the directions-of-arrival of the impinging wavefronts. The signal model is given by

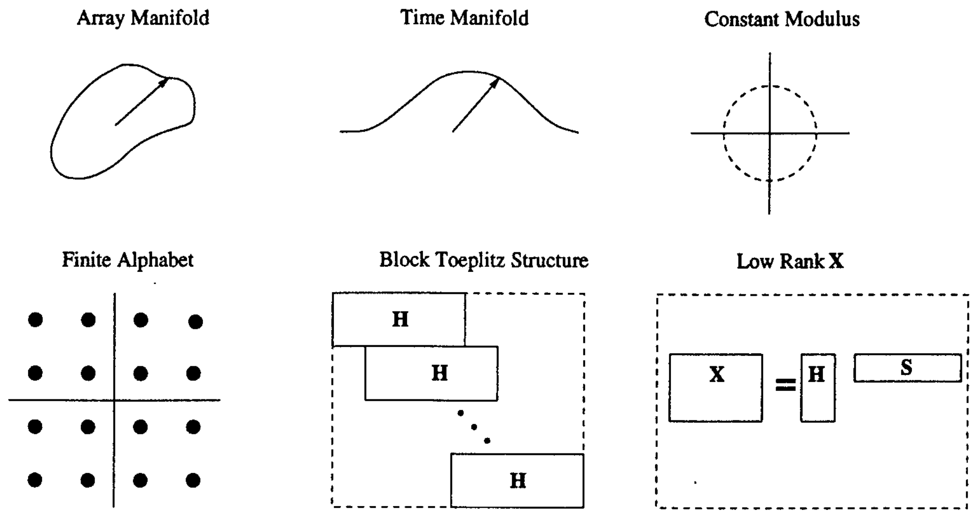

Figure 2: Space-time structures

$$\mathbf{x}(t) \;=\; \mathbf{A}\mathbf{u}(t) + \mathbf{n}(t) \qquad (6)$$

where $\mathbf{A}$ is an $m \times Q$ matrix whose columns are the array response vectors for each wavefront (assuming no multipath)

$$\mathbf{A} \;=\; \left[\; \mathbf{a}(\theta_1) \quad \cdots \quad \mathbf{a}(\theta_Q) \;\right] ,$$

$\mathbf{u}(t)$ contains the fading signals from the Q users

$$\mathbf{u}(t) \;=\; [\alpha_1(t)u_1(t-\tau_1) \;\cdots\; \alpha_l(t)u_Q(t-\tau_Q)]^T$$

and

$$u_q(t) \;=\; \sum_k s_q(k)g(t-kT)$$

The sampled block signal model then takes the following form

$$\mathbf{X} \;=\; \mathbf{A}\;\mathbf{S} \;+\; \mathbf{N} \qquad (7)$$

In the subspace approach, we seek to estimate $\mathbf{A}$ from the array data by exploiting the underlying array manifold structure. When the number of antennas m is greater than the

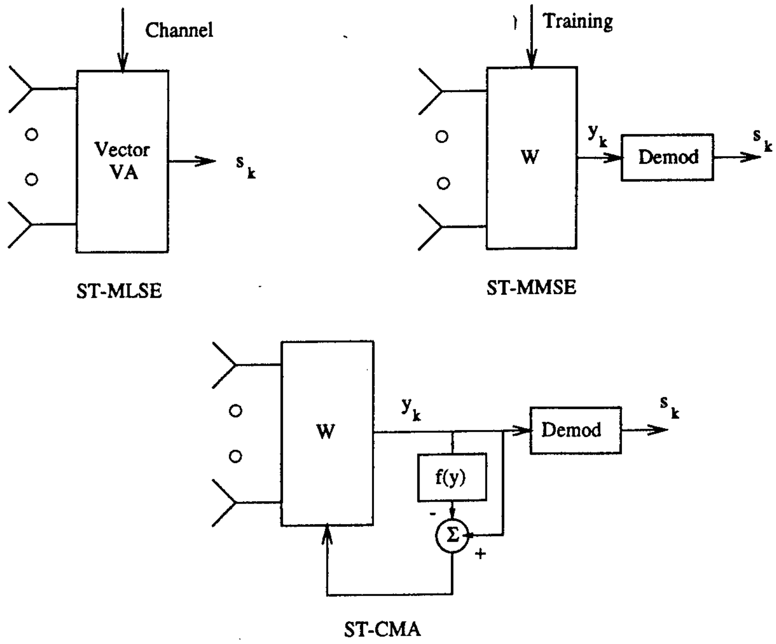

Figure 3: Different structures for space-time processing

number of signals Q, the signal $\mathbf{x}(t)$ in the absence of noise is confined to a subspace, referred to as the *signal subspace.*

We first estimate this signal subspace from the received data $\mathbf{X}$. We then search for an $m \times Q$ matrix $\mathbf{A}$ whose *columns lie on the array manifold* and whose (column) subspace matches the estimated signal subspace. A good estimate of the signal subspace is given by the first Q dominant eigenvectors of the space-only $m \times m$ covariance matrix $\mathbf{R}_{xx} = E(\mathbf{x}\mathbf{x}^H)$. If $\mathbf{E}_s$ is a matrix of these eigenvectors, then the subspace fitting approach estimates $\mathbf{A}$ to minimize the following criterion

$$\min_{\mathbf{A}} \quad \|\mathbf{E}_s - \mathbf{A}\mathbf{Z}\|_F^2$$

where $\mathbf{Z}$ is an arbitrary $Q \times Q$ square matrix.

Once $\mathbf{A}$ is estimated, we have the array vector for the desired signal. The MMSE and ML estimators (assuming no multipath) of $u_q(t)$ are identical and are given by

$$\mathbf{w}_q \;=\; \mathbf{R}_{xx}^{-1}\mathbf{a}(\theta_q) \tag{8}$$

$\mathbf{w}_q$ is a (space-only) beamformer that has been studied extensively.

When multipath and delay spread is present, the solution in Eq. (8) will have a poor or even disastrous performance and improved techniques are needed. If we use a space-time MMSE (ST-MMSE) structure, we can extend the above subspace methods to compute the optimum beamformer.

Temporal structure methods These techniques include a vast range that spans from the well studied CM and HOS methods to the more recent second order methods that exploit the cyclostationarity of the received signal.

The fading and dynamics of the mobile propagation channel create special problems for blind techniques, and their performance in mobile channels is only recently gaining attention. A widely known class of simple blind algorithms is the so-called Bussgang class that contains among others the CM 1-2, CM 2-2, Sato, and Decision-Directed (DD) algorithms.

The use of the CM algorithms in spatial processing has been studied since the 1980s (see Figure 2 for the structure of the CM). They are applicable to all linear (QAM) modulation schemes and have been successfully used in practice for several years. It was only recently shown [6] that any sub-Gaussian constellation (such as QAM or other wireless modulation formats) can be blindly retrieved using the constant modulus criteria. Most results to date have been for channels without delay spread. In the presence of delay spread, we need a space-time version of CM. We describe the ST-CM 2-2 algorithm, which is a good prototype for a range of blind space-time methods.

Contrary to non-blind techniques where a training signal drives the recursive algorithms, in the CM approach we replace the training signals by a modulus corrected version of the output signal. The CM 2-2 minimizes the following cost function

$$\min_{W} J(W) \;=\; E\left||y(k)|^2 - 1\right|^2 \tag{9}$$

where $y(k)$ is the output of the space-time filter.

The resulting LMS-type algorithm is given by

$$W(k+1) \;=\; W(k) \;-\; \mu\, X^*(k)\, y(k)\, \Big(|y(k)|^2 \;-\; 1\Big). \tag{10}$$

$W(k+1)$, under the right conditions, approaches the optimum ST-MMSE solution.

Important performance issues for blind algorithms are speed of convergence, ability to reach the global optimum solution, and capacity to track time varying mobile channels. Several refinements that improve these parameters of the CM 2-2 have been recently reported in the literature. These include the Least-Squares CM [7], the Normalized CM [8], the Normalized Sliding Window CM [9], Normalized CM with Relaxation [10], and Decision feedback CM [11], [12], [13], [14]. A recent survey of CM algorithms can be found in [15]. The application of CM algorithms to space-time processing is an emerging field.

Polyphase methods Following the path-breaking paper by Tong, Xu, and Kailath [16] that presented a blind channel identification method using oversampling and relied only on second order statistics, a number of techniques that exploit cyclostationarity have since dominated the blind-deconvolution literature.

Polyphase methods provide a blind solution by starting with the data

$$\mathbf{X}(k) = \mathbf{H}\mathbf{S}(k) + \mathbf{N}(k) \tag{11}$$

or its second order statistics. They then extract $\mathbf{H}$ and $\mathbf{S}$ by exploiting the tallness structure (obtained via oversampling) of $\mathbf{H}$. See [17] for a tutorial presentation of polyphase techniques.

4 Simulation examples

A TDMA (IS-54) system model with a typical urban (TU) channel [18] and mobile speeds of 50 and 100 km/h are assumed. The desired signal has an angle spread of 10^o and arrives from a mean direction of 0^o referenced to the antenna broadside, whereas the two interfering signals have an angle spread of 3^o each and arrive from mean directions of 30^o and -30^o, respectively. Each interferer is 7 dB below the signal. A four-element uniform linear array with $\frac{\lambda}{2}$ spacing is used. The space-time processor uses two time taps fractionally spaced at $T/2$ per antenna. Two $T/2$ taps are adequate given the relatively small delay spread in a TU channel. The ST-MMSE coefficients are initialized with a least-squares solution based on the 14-symbol training sequence and are then updated through the TDMA time slot (162 symbols) using a blind adaptive algorithm (RLS - decision directed (see [9])) with a

forgetting factor of $\lambda = 0.9$.

Figure 4 plots the BER performance of four simulation experiments at various carrier-to-noise ratios (CNRs). Two plots are shown for the single antenna conventional differential demodulator (no spatial processing) at the two mobile speeds. Note the poor performance even at high CNR, indicating an interference-limited situation. The results for the ST-MMSE show significant improvement demonstrating the value of space-time processing in real environments.

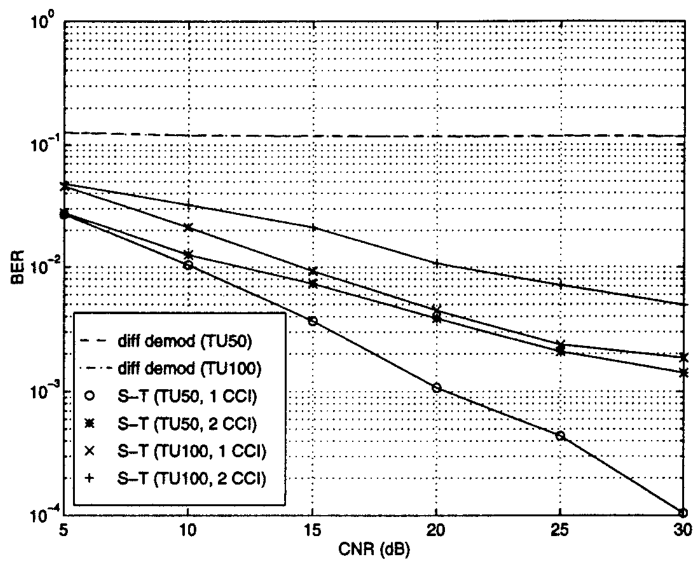

Figure 4: ST-MMSE: Interference Reduction Performance

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